

METHOD AND ARRANGEMENT FOR NOISE VARIANCE AND SIR
ESTIMATION

5 **Field of the Invention**

This invention relates to noise variance and
signal/interference ratio (SIR) estimation, and
particularly though not exclusively to such estimation in
10 wireless communication receivers. It will be understood
that, as used herein, the terms 'noise' and
'interference' are to be considered synonymous, with each
encompassing both noise and interference.

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Background of the Invention

In the field of this invention it is known that many
parts of a wireless communications receiver often require
20 an estimation of noise variance and/or SIR. This is
needed for purposes of power control, threshold
determination for various algorithms, quantisation of
soft-decision information for channel decoding purposes
to name but a few.

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For BPSK (Binary Phase Shift Key) and QPSK (Quadrature
Phase Shift Key) modulation the conventional method for
estimating the SIR at the output of a detector relies on
estimating output noise variance using the following
30 equality known (for example) from the publication by

Papoulis and Pillai, entitled 'Probability, Random Variables and Stochastic Processes', 3rd Ed. 1991,

$$\hat{\sigma}_z^2 = E\left(\left|\hat{d}_n^{(k)}\right|^2\right) - E\left(\left|\hat{d}_n^{(k)}\right|\right)^2$$

where $\hat{\sigma}_z^2$ represents variance, E represents mean value

5 and $\hat{d}_n^{(k)}$ represents the detector output.

This yields the following result:

$$SIR^{(k)} = \frac{P^{(k)} - \hat{\sigma}_z^2}{\hat{\sigma}_z^2} = \frac{E\left(\left|\hat{d}_n^{(k)}\right|\right)^2}{E\left(\left|\hat{d}_n^{(k)}\right|^2\right) - E\left(\left|\hat{d}_n^{(k)}\right|\right)^2}$$

where SIR represents the SIR of the k^{th} sequence at the
10 detector output, and $P^{(k)}$ represents the average power of
the k^{th} sequence at the detector output.

However, this approach has the disadvantage(s) that the
accuracy of this method at low SIR is poor since it
15 suffers from a bias term. An analysis of the bias term
and a correction method has been suggested in GB patent
application no. 0128475.1. However, the suggested
correction method requires a look-up table to correct for
the aforementioned problem, and the estimation variance
20 is also increased when correcting the bias.

A need therefore exists for a method and arrangement for
noise variance and SIR estimation wherein the
abovementioned disadvantage(s) may be alleviated.

Statement of Invention

In accordance with a first aspect of the present invention there is provided a method for noise variance
5 estimation as claimed in claim 1.

In accordance with a second aspect of the present invention there is provided an arrangement for noise variance estimation as claimed in claim 10.

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Preferably, the second noise variance signal is produced by applying to the first noise variance signal a function substantially equal to the detector's transfer function.

15 Preferably, the first noise variance signal is derived from a midamble portion of the received signal.

Preferably, an estimate of total power at the detector output is produced from the second noise variance signal
20 and an SIR signal representative of SIR in the received signal.

Brief Description of the Drawings

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One method and arrangement for noise variance and SIR estimation incorporating the present invention will now be described, by way of example only, with reference to the accompanying drawings, in which:

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FIG. 1 shows a block schematic diagram illustrating a 3GPP radio communication system in which the present invention may be used;

5 FIG. 2 shows a block schematic diagram illustrating a generic burst structure employed in the system of FIG. 1; and

10 FIG. 3 shows a block schematic diagram illustrating a detector arrangement incorporating a noise variance and SIR estimator utilising the present invention.

15 Description of Preferred Embodiment(s)

The following preferred embodiment of the present invention will be described in the context of a 3GPP (3rd Generation Partnership Project) UMTS (Universal Mobile Telecommunication System) Radio Access Network (UTRAN) system operating in TDD mode. Referring firstly to FIG. 1, a typical, standard UTRAN system 100 is conveniently considered as comprising: a terminal/user equipment domain 110; a UMTS Terrestrial Radio Access Network domain 120; and a Core Network domain 130.

In the terminal/user equipment domain 110, terminal equipment (TE) 112 is connected to mobile equipment (ME) 114 via the wired or wireless *R* interface. The ME 114 is also connected to a user service identity module (USIM) 116; the ME 114 and the USIM 116 together are considered

as a user equipment (UE) 118. The UE 118 communicates data with a Node B (base station) 122 in the radio access network domain 120 via the wireless *Uu* interface. Within the radio access network domain 120, the Node B 122
5 communicates with a radio network controller (RNC) 124 via the *Iub* interface. The RNC 124 communicates with other RNC's (not shown) via the *Iur* interface. The Node B 122 and the RNC 124 together form the UTRAN 126. The RNC 124 communicates with a serving GPRS service node (SGSN) 132
10 in the core network domain 130 via the *Iu* interface. Within the core network domain 130, the SGSN 132 communicates with a gateway GPRS support node (GGSN) 134 via the *Gn* interface; the SGSN 132 and the GGSN 134 communicate with a home location register (HLR) server
15 136 via the *Gr* interface and the *Gc* interface respectively. The GGSN 134 communicates with public data network 138 via the *Gi* interface.

Thus, the elements RNC 124, SGSN 132 and GGSN 134 are
20 conventionally provided as discrete and separate units (on their own respective software/hardware platforms) divided across the radio access network domain 120 and the core network domain 130, as shown the FIG. 2.

25 The RNC 124 is the UTRAN element responsible for the control and allocation of resources for numerous Node B's 122; typically 50 to 100 Node B's may be controlled by one RNC. The RNC also provides reliable delivery of user traffic over the air interfaces. RNC's communicate with
30 each other via the *Iur* interface.

The SGSN 132 is the UMTS Core Network element responsible for Session Control and interface to the HLR. The SGSN keeps track of the location of an individual UE and performs security functions and access control. The SGSN
5 is a large centralised controller for many RNCs.

The GGSN 134 is the UMTS Core Network element responsible for concentrating and tunnelling user data within the core packet network to the ultimate destination (e.g.,
10 internet service provider - ISP).

Such a UTRAN system and its operation are described more fully in the 3GPP technical specification documents 3GPP TS 25.401, 3GPP TS 23.060, and related documents,
15 available from the 3GPP website at www.3gpp.org, and need not be described in more detail herein.

The physical layer of UTRA TDD mode provides physical channels that carry transport channels from the MAC
20 (Medium Access Control) sub-layer of UMTS Layer 2. A physical channel is defined by frequency, timeslot, channelisation code, burst type, and radio frame allocation. In UMTS Layer 2, in each time slot, three burst structures (as shown generically in FIG. 2) can be
25 supported, each consisting of two data fields (210, 230), a midamble (220) and a guard period (240).

The data fields contain the data symbols from the transport channels, after the processes of coding,
30 multiplexing, interleaving, and modulation. The midamble field contains the training sequence, which is used in a

number of Layer 1 algorithms, such as channel estimation. The guard period, GP, is used to accommodate any timing inaccuracies, from propagation delays, channel dispersion, and power ramping in the transmitter. The
 5 different burst types and their associated field lengths in chips are given in the table below:

Burst Type	Data Field 1	Data Field 2	Midamble L_m	GP
Burst Type 1	976	976	512	96
Burst Type 2	1104	1104	256	96
Burst Type 3	976	880	512	192

The received sequence in the data payload areas of the
 10 burst is given by

$$\mathbf{e} = \mathbf{A}\mathbf{d} + \mathbf{n}$$

where

$$\mathbf{d} = (\mathbf{d}^{(1)T}, \mathbf{d}^{(2)T}, \dots, \mathbf{d}^{(K)T})^T = (d_1, d_2, \dots, d_{KN})^T,$$

$(\cdot)^T$ denotes transposition, K is the number of data
 15 sequences $k=1, \dots, K$, and N is the number of symbols per data sequence $n=1, \dots, N$. The data sequence for the k^{th} user is given by $\mathbf{d}^{(k)} = (d_1^{(k)}, d_2^{(k)}, \dots, d_N^{(k)})^T$. The noise sequence \mathbf{n} , denoted by $\mathbf{n} = (n_1, n_2, \dots, n_{NQ+W-1})^T$, has zero mean and covariance matrix $\mathbf{R}_n = E(\mathbf{n}\mathbf{n}^H)$, where $(\cdot)^H$ denotes conjugate
 20 transposition. The matrix has dimensions $(NQ+W-1) \times KN$ and the elements are given by:

$$A_{Q(n-1)+l, N(k-1)+n} = \begin{cases} \mathbf{b}^{(k)} & k=1 \dots K, l=1 \dots Q+W-1, n=1 \dots N \\ 0 & \text{else} \end{cases}$$

where $\mathbf{b}^{(k)} = \mathbf{h}^{(k)} * \mathbf{c}^{(k)}$, $\mathbf{h}^{(k)}$ is the impulse response of the k^{th} user, W is the length of the channel impulse response, $*$ denotes discrete time convolution, $\mathbf{c}^{(k)}$ is the spreading code of the k^{th} user and Q is the length of the spreading sequence in chips.

The output of the detector is given by

$$\hat{\mathbf{d}} = f(\mathbf{e}) = \mathbf{r} + \mathbf{z}$$

where $f(\cdot)$ denotes the transfer function of the detector, the vector \mathbf{r} contains the desired symbols, and the vector \mathbf{z} contains noise plus interference. The average power for the k^{th} sequence at the output of the detector is given by

$$P^{(k)} = E\left(\left|\hat{d}_n^{(k)}\right|^2\right)$$

Expanding $P^{(k)}$ produces

$$P^{(k)} = E\left(\left|r_{n+(k-1)Q}\right|^2\right) + E\left(r_{n+(k-1)Q} z_{n+(k-1)Q}^*\right) + E\left(r_{n+(k-1)Q}^* z_{n+(k-1)Q}\right) + E\left(\left|z_{n+(k-1)Q}\right|^2\right)$$

Under the assumption that the noise is uncorrelated with the signal vector \mathbf{r} , the average power for the k^{th} sequence becomes

$$P^{(k)} = E\left(\left|r_{n+(k-1)Q}\right|^2\right) + \sigma_z^2$$

where $E(\cdot)$ is the statistical average, $\sigma_z^2 = E\left(\left|z_{n+(k-1)Q}\right|^2\right)$ is the noise variance at the output of the detector, and

$E\left(\left|r_{n+(k-1)Q}\right|^2\right)$ is the signal power for the k^{th} sequence. The

SIR at the output of the detector for the k^{th} sequence is therefore given by

$$SIR^{(k)} = \frac{P^{(k)} - \sigma_z^2}{\sigma_z^2} = \frac{E(|r_{n+(k-1)Q}|)}{\sigma_z^2}$$

The conventional method for estimating the SIR relies on estimating the detector output noise variance using the

5 following equality mentioned above:

$$\hat{\sigma}_z^2 = E(|\hat{d}_n^{(k)}|^2) - E(|\hat{d}_n^{(k)}|)^2$$

to yield the following result:

$$SIR^{(k)} = \frac{P^{(k)} - \hat{\sigma}_z^2}{\hat{\sigma}_z^2} = \frac{E(|\hat{d}_n^{(k)}|)^2}{E(|\hat{d}_n^{(k)}|^2) - E(|\hat{d}_n^{(k)}|)^2}$$

10 As discussed above, the accuracy of this approach at low SIR is poor since it suffers from a bias term, which may be corrected by use of a look-up table.

As will be discussed in greater detail below, the
15 following preferred embodiments of the present invention do not suffer from such a bias term and therefore do not require a look-up table to correct for the aforementioned problem.

20 Referring now to FIG. 3, in detector arrangement 300 (which may be used in Node B 122 or User Equipment 118) a new technique for estimating the SIR at the output of a detector 310 uses an estimate of the detector input noise variance (derived in channel estimator 320) to provide an
25 estimate of the detector output noise variance (derived in SIR estimator 330). In the channel estimator 320 the detector input noise variance is derived from the

midamble portion of the burst using locally reconstructed replicas of the transmitted signal which are removed from the received set of samples providing a residual term that contains the noise of the channel. In the SIR

5 estimator 330, by deriving the transfer function of the detector an estimate of the detector output noise variance is estimated.

The estimated output noise variance then allows an improved estimate of the SIR ($SIR^{(1)} \dots SIR^{(K)}$) at the detected output. Typically, the SIR at the output of the detector is used for deriving soft decision quantisation levels for application to channel decoding algorithms.

15 In the following description, two types of CDMA (Code Division Multiple Access) detector are considered, namely single user detector (SUD) and multiuser detector (MUD). It will be understood that the invention is also applicable to other types of detector such as a RAKE receiver.

The technique described here is based on first estimating the noise variance at the input to the detector and then mapping the input noise variance to the output noise variance using the transfer function of the detector.

The process of estimating the noise variance at the input to the detector is carried out using the midamble portion of the burst. Considering the received sequence of chip spaced samples $\mathbf{e} = (e_1, e_2, \dots, e_{L_b})^T$, where L_b is the burst

length, a locally reconstructed version for the midamble portion of the burst is defined as follows:

$$\bar{\mathbf{e}} = (\bar{e}_1, \bar{e}_2, \dots, \bar{e}_{L_m+W-1})^T = \left(\sum_k (\mathbf{h}^{(k)} * \mathbf{m}^{(k)}) \right)^T$$

where $\bar{\mathbf{e}}$ denotes locally reconstructed version and $\mathbf{m}^{(k)}$ is the midamble sequence of the k^{th} user. The noise variance estimate at the input to the detector is given by

$$\hat{\sigma}^2 = E \left(\left((e_W, e_{W+1}, \dots, e_X) - (\bar{e}_W, \bar{e}_{W+1}, \dots, \bar{e}_X) \right)^2 \right)$$

where $X \leq L_m$ and the starting position is W since the first $W-1$ samples from the midamble portion of the burst are corrupted by the data portion of the burst.

Multiuser Detection

Under the assumption that the noise is white with variance σ^2 , the Minimum Mean Squared Error (MMSE) block linear equalizer solution to symbol estimation is given (as known from the publication of Klein, Kaleh and Baier entitled 'Zero Forcing and Minimum Mean-Square-Error Equalization for Mutliuser Detection in Code-Division Multiple-Access Channels' in IEEE Trans VT, VOL.45, No.2, May 1996, pp276-287) by

$$\hat{\mathbf{d}} = (\mathbf{A}^H \mathbf{A} + \sigma^2 \mathbf{I})^{-1} \mathbf{A}^H \mathbf{e} = f(\mathbf{e}) = \mathbf{r} + \mathbf{z}$$

where \mathbf{I} is the identity matrix and

$$\mathbf{r} = (\mathbf{A}^H \mathbf{A} + \sigma^2 \mathbf{I})^{-1} \mathbf{A}^H \mathbf{A} \mathbf{d} = (r_1, r_2, \dots, r_{KN})^T \quad (1)$$

$$\mathbf{z} = (\mathbf{A}^H \mathbf{A} + \sigma^2 \mathbf{I})^{-1} \mathbf{A}^H \mathbf{n} = (z_1, z_2, \dots, z_{KN})^T \quad (2)$$

From equation (2), the noise variance seen at the output of the MUD is given by

$$\sigma_z^2 = E\left(|z_{n+(k-1)Q}|^2\right) = \left(\|\mathbf{b}^{(k)}\|^2 + \sigma^2\right)^{-2} \|\mathbf{b}^{(k)}\|^2 \sigma^2$$

where $\|\cdot\|$ denotes vector norm, and σ^2 represents the noise variance at the input of the MUD. By replacing σ^2 with the estimate of the MUD input noise variance $\hat{\sigma}^2$, we have a direct method for estimating the MUD output noise variance $\hat{\sigma}_z^2$. For completeness, the estimate $\hat{\sigma}_z^2$ of the MUD output noise variance can be written as

$$\hat{\sigma}_z^2 = f(\hat{\sigma}^2) = \left(\|\mathbf{b}^{(k)}\|^2 + \hat{\sigma}^2\right)^{-2} \|\mathbf{b}^{(k)}\|^2 \hat{\sigma}^2$$

where $f(\hat{\sigma}^2)$ represents the noise transfer function of the detector.

Using the new estimate for the output noise variance, the SIR at the output of the MUD for the k^{th} sequence is defined by

$$SIR^{(k)} = \frac{E\left(|r_{n+(k-1)Q}|^2\right) + \delta(\hat{\sigma}^2)}{\hat{\sigma}_z^2}$$

where the error term $\delta(\hat{\sigma}^2)$ is given by **BEST AVAILABLE COPY**

$$\delta(\hat{\sigma}^2) = \|\mathbf{b}^{(k)}\|^2 \left(\frac{\sigma^2}{\left(\|\mathbf{b}^{(k)}\|^2 + \sigma^2\right)^2} - \frac{\hat{\sigma}^2}{\left(\|\mathbf{b}^{(k)}\|^2 + \hat{\sigma}^2\right)^2} \right)$$

20

It is clear from the above set of equations that when $\hat{\sigma}^2 = \sigma^2$ we have the following

$$SIR^{(k)} = \frac{E(|r_{n+(k-1)Q}|^2)}{\sigma_z^2}$$

It will therefore be understood that the accuracy of the above technique is directly related to the quality of the noise variance estimate, $\hat{\sigma}^2$, at the input of the MUD.

Single User Detection

For the single user detector case the received sequence is written as

$$\mathbf{e} = \mathbf{A}\mathbf{d} + \mathbf{n} = \mathbf{H}\mathbf{C}\mathbf{d} + \mathbf{n}$$

The matrix \mathbf{H} has dimensions $(NQ+W-1) \times NQ$ and its elements are given by

$$H_{i,v} = \begin{cases} h_{i-v+1} & 1 \leq i-v+1 \leq W \\ 0 & \text{else} \end{cases}$$

where $\mathbf{h} = (h_1, h_2, \dots, h_W)^T$, $i = 1, \dots, NQ+W-1$, and $v = 1, \dots, NQ$. The matrix \mathbf{C} has dimensions $NQ \times KN$ and its elements are given by

$$C_{Q(n-1)+q, N(k-1)+n} = \begin{cases} c_q^{(k)} & q = 1, \dots, Q, n = 1, \dots, N, k = 1, \dots, K \\ 0 & \text{else} \end{cases}$$

For Minimum Mean Squared Error (MMSE) symbol estimation and under the assumption that the noise is white with variance σ^2 , the output of the SUD is given by

$$\hat{\mathbf{d}} = \mathbf{C}^H (\mathbf{H}^H \mathbf{H} + \sigma^2 \mathbf{I})^{-1} \mathbf{H}^H \mathbf{e} = \mathbf{f}(\mathbf{e}) = \mathbf{r} + \mathbf{z}$$

where

$$\mathbf{r} = \mathbf{C}^H (\mathbf{H}^H \mathbf{H} + \sigma^2 \mathbf{I})^{-1} \mathbf{H}^H \mathbf{H} \mathbf{C} \mathbf{d} = (r_1, r_2, \dots, r_{KN})^T \quad (3)$$

$$\mathbf{z} = \mathbf{C}^H (\mathbf{H}^H \mathbf{H} + \sigma^2 \mathbf{I})^{-1} \mathbf{H}^H \mathbf{n} = (z_1, z_2, \dots, z_{KN})^T \quad (4)$$

From equation (4), the noise variance seen at the output of the SUD is given by

$$\sigma_z^2 = E\left(|z_{n+(k-1)Q}|^2\right) = G \times (\|\mathbf{h}\|^2 + \sigma^2)^2 \|\mathbf{h}\|^2 \sigma^2 \quad \dots \dots \dots$$

where $\|\cdot\|$ denotes vector norm, the multiplier G comes from the matrix \mathbf{C} , and in general $G = \|\mathbf{c}^{(k)}\|^2 = Q$.

By replacing σ^2 with the estimate of the SUD input noise variance $\hat{\sigma}^2$, we have a direct method for estimating the SUD output noise variance σ_z^2 . For completeness, the estimate $\hat{\sigma}_z^2$ of the SUD output noise variance can be written as

$$\hat{\sigma}_z^2 = f(\hat{\sigma}^2) = Q \times (\|\mathbf{h}\|^2 + \hat{\sigma}^2)^2 \|\mathbf{h}\|^2 \hat{\sigma}^2$$

where G is replaced with Q and $f(\hat{\sigma}^2)$ is the noise transfer function of the detector. Using the new estimate for the output noise variance, the SIR at the output of the SUD for the k^{th} sequence is defined by

$$SIR^{(k)} = \frac{E\left(|r_{n+(k-1)Q}|^2\right) + \delta(\hat{\sigma}^2)}{\hat{\sigma}_z^2}$$

where the error term $\delta(\hat{\sigma}^2)$ is given by

$$\delta(\hat{\sigma}^2) = Q \times \|\mathbf{h}\|^2 \left(\frac{\sigma^2}{(\|\mathbf{h}\|^2 + \sigma^2)^2} - \frac{\hat{\sigma}^2}{(\|\mathbf{h}\|^2 + \hat{\sigma}^2)^2} \right)$$

It is clear from the above set of equations that when $\hat{\sigma}^2 = \sigma^2$ we have the following

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$$SIR^{(k)} = \frac{E(|r_{n+(k-1)Q}|^2)}{\sigma_z^2}$$

It will therefore be understood that the accuracy of the above technique is directly related to the quality of the noise variance estimate, $\hat{\sigma}^2$, at the input of the SUD.

It will be appreciated that the method described above for noise variance and SNIR estimation may be carried out in software running on a processor (not shown - e.g., in User Equipment such as 118 or a Node B such as 122), and that the software may be provided as a computer program element carried on any suitable data carrier (also not shown) such as a magnetic or optical computer disc.

It will be also be appreciated that the arrangement described above for noise variance and SNIR estimation may alternatively be carried out in hardware, for example in the form of an integrated circuit (not shown) such as an FPGA (Field Programmable Gate Array) or ASIC (Application Specific Integrated Integrated Circuit).

It will be understood that the method and arrangement for noise variance and SIR estimation described above provides the following advantages that the accuracy of this technique is not poor at low SIR, since it does not suffer from a bias term, nor does it require correction therefor using a look-up table. An additional advantage is that any increase in estimation variance resulting from bias correction may be avoided.